BICM-ID scheme for clipped DCO-OFDM in visible light communications

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Abstract: Visible light communication (VLC) is recommended for indoor transmissions in 5G network, whereby DC-biased optical orthogonal frequency division multiplexing (DCO-OFDM) is adopted to eliminate the inter-symbol interference (ISI) but suffers from considerable performance loss induced by clipping distortion. In this paper, bit-interleaved coded modulation with iterative demapping and decoding (BICM-ID) scheme for clipped DCO-OFDM is investigated to enhance the performance of VLC systems. In order to further mitigate the clipping distortions, a novel soft demapping criterion is proposed, and a simplified demapping algorithm is developed to reduce the complexity of the proposed criterion. Simulation results illustrate that the enhanced demapping algorithm achieves a significant performance gain.

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1. Introduction

Recently, visible light communication (VLC) is regarded as a promising candidate technology to provide high-rate Internet access for indoor scenarios in 5G wireless communication network [1]. In VLC systems, the transmitted signals are directly used to drive on light emitting diodes (LED) and have to be real-valued and non-negative for intensity modulation and direct detection (IM/DD) [2]. Optical orthogonal frequency-division multiplexing (OFDM) is commonly adopted in VLC systems due to its strong resistance against inter-symbol interference (ISI) [3].

Several optical OFDM schemes [4, 5] are proposed to satisfy the real-valued and positive constrains of VLC systems with careful tradeoff between power and spectral efficiency. For DC-biased OFDM (DCO-OFDM) [4], Hermitian symmetry is applied in frequency domain and additional DC bias guarantees the positive time-domain outputs. Since the electronic to optical power transfer characteristic of LED exhibits an intense non-linearity, DCO-OFDM signals should be clipped based on dynamic range of LEDs. Due to the inherent high peak-to-average power ratio (PAPR) of DCO-OFDM, clipping operation results in a considerable distortion in frequency domain, which deteriorates the performance inevitably [6]. Specific PAPR or cubic metric (CM) reduction methods for DCO-OFDM are proposed in [7–9] to achieve significant clipping distortion mitigation, whereas these algorithms have high complexity and are difficult to be implemented. In contrast, [10] proposes a low-complexity adaptive scaling and biasing scheme for DCO-OFDM with limited performance gain. Besides that, low-complexity maximum likelihood sequence detection for clipped DCO-OFDM signals is capable of approaching the performance of ideal case [11].

Bit-interleaved coded modulation with iterative demapping and decoding (BICM-ID) is a preferred option for VLC systems, which adopts bitwise interleaving to increase the diversity order and provides strong error correction capability with soft information feedback at the receiver [12]. In this paper, BICM-ID scheme for clipped DCO-OFDM is investigated for VLC systems to enhance the performance. The conventional demapper for clipped DCO-OFDM neglects the correlation between clipping distortions over different subcarriers in frequency domain and suffers a significant performance loss [6]. Considering the characteristic of clipped DCO-OFDM, a novel demapping extrinsic log likelihood ratio (LLR) criterion is derived by incorporating the sequence likelihood, which provides more accurate soft information for the decoder to improve error correction. Besides that, a simplified demapping algorithm is developed to reduce the high complexity of the proposed LLR calculation. Simulation results verify that BICM-ID scheme with the proposed soft demapper achieves a significant performance gain and clipping distortion mitigation in VLC systems.

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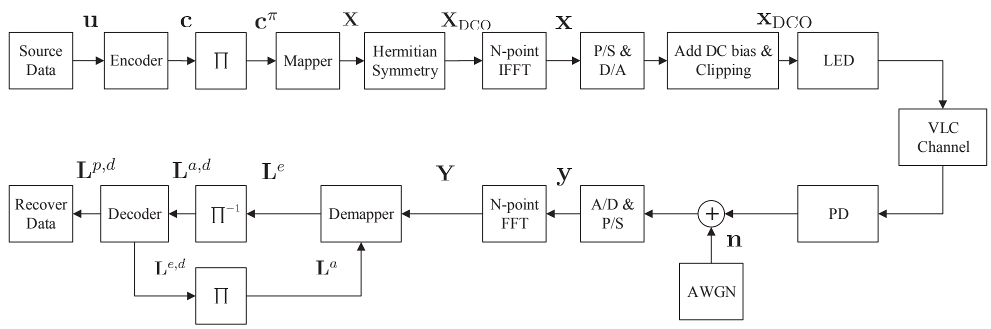

Fig. 1. The block diagram of DCO-OFDM transceiver using BICM-ID technology.

2. System model

In this section, BICM-ID scheme with the conventional demapper for clipped DCO-OFDM is proposed as depicted in Fig. 1. At the transmitter, The source bits $\mathbf{u}$ are converted into $\mathbf{c}$ and $\mathbf{c}^\pi$ after the encoder and the interleaver. The mapper transfers the encoded bits into symbol vector $\mathbf{X} = \left[X_1, X_2, ..., X_{N_{used}}\right]^{\mathrm{T}} \in \mathscr{X}^{N_{used}}$ in each N-point DCO-OFDM block according to the 2^m-ary constellation set $\mathscr{X}$, where N_{used} denotes the number of occupied subcarriers. $\mathbf{X}_{\mathrm{DCO}}$ is generated in Hermitian symmetry, which is defined as

$$\mathbf{X}_{\mathrm{DCO}} = HS(\mathbf{X}) = \left[0, X_1, ..., X_{N_{used}}, 0, ..., 0, X^*_{N_{used}}, ..., X^*_1\right]^{\mathrm{T}}. \tag{1}$$

$\mathbf{X}_{\mathrm{DCO}}$ is fed to inverse fast Fourier transformation (IFFT) converter to generate real-valued time-domain signal $\mathbf{x}$, and we have

$$x_n = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X_{\mathrm{DCO},k} \exp\left(j\frac{2\pi kn}{N}\right), n = 0, 1, .., N-1, \tag{2}$$

where x_n and $X_{\mathrm{DCO},k}$ are the n-th element of $\mathbf{x}$ and the k-th element of $\mathbf{X}_{\mathrm{DCO}}$, respectively. Since LED is driven by the positive electronic signal, DC bias B_{DC} should be added to $\mathbf{x}$. Considering the nonlinear LED transfer characteristic, $\mathbf{x} + B_{\mathrm{DC}}$ is clipped beyond the LED dynamic region of $[A_{\min}, A_{\max}]$ [13]. The transmitted DCO-OFDM signals $\mathbf{x}_{\mathrm{DCO}}$ could be expressed as

$$\mathbf{x}_{\mathrm{DCO}} = \lfloor \mathbf{x} + B_{\mathrm{DC}} \rceil, \tag{3}$$

where the double-side clipping $\lfloor\cdot\rceil$ is defined as

$$\lfloor z \rceil = \begin{cases} A_{\min},\ z < A_{\min} \\ A_{\max},\ z > A_{\max} \\ z,\ \ else \end{cases}. \tag{4}$$

DCO-OFDM signals are modulated on the illumination of the emitted visible light and transmitted over VLC channel, which is assumed as additive white Gaussian noise (AWGN) channel [6]. At the receiver, the photodiode (PD) component transfers the optical signals into electronic format $\mathbf{y} = \mathbf{x}_{\mathrm{DCO}} + \mathbf{n}$, where $\mathbf{n}$ denotes the N-dimension independent and identical Gaussian distributed noise with zero mean and variance of σ^2. Traditionally, clipping distortions of DCO-OFDM signals are modeled as non-correlated noise, the conventional iterative demapper is simply applied with the received frequency-domain symbol vector $\mathbf{Y}$ after FFT

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converter. The demapper of BICM-ID scheme exports the extrinsic LLR as input soft information of the decoder iteratively. In terms of tradeoff between complexity and performance, max-sum approximation of the maximum *a posteriori* probability in log-domain (Max-Log-MAP) algorithm [14] is invoked to calculate the extrinsic LLR. $\mathbf{L}_k^a$ represents the m-length *a priori* LLR vector of the symbol on the k-th subcarrier and the total *a priori* LLR vector $\mathbf{L}^a$ is grouped as $\mathbf{L}^a = \left[(\mathbf{L}_1^a)^{\mathrm{T}}, (\mathbf{L}_2^a)^{\mathrm{T}}, ..., \left(\mathbf{L}_{N_{used}}^a\right)^{\mathrm{T}}\right]^{\mathrm{T}}$. Assuming that Y_k is the k-th element of $\mathbf{Y}$, the extrinsic LLR of the i-th bit on the k-th subcarrier $L_{k,i}^a$ is acquired as [15]

$$\begin{aligned} L_{k,i}^e &\approx \log \frac{\max\limits_{\overline{X}_k \in \mathscr{X}_i^{(0)}} \left\{ P\left(Y_k|\overline{X}_k\right) P\left(\overline{X}_k|\mathbf{L}_k^a\right)\right\}}{\max\limits_{\overline{X}_k \in \mathscr{X}_i^{(1)}} \left\{ P\left(Y_k|\overline{X}_k\right) P\left(\overline{X}_k|\mathbf{L}_k^a\right)\right\}} - L_{k,i}^a \\ &= \max_{\overline{X}_k \in \mathscr{X}_i^{(0)}} \left\{ -\frac{1}{\sigma^2}\left\|\overline{X}_k - Y_k\right\|^2 + \frac{1}{2}\mathbf{s}^{\mathrm{T}}\left(\overline{X}_k\right)\mathbf{L}_k^a \right\} - \max_{\overline{X}_k \in \mathscr{X}_i^{(1)}} \left\{ -\frac{1}{\sigma^2}\left\|\overline{X}_k - Y_k\right\|^2 + \frac{1}{2}\mathbf{s}^{\mathrm{T}}\left(\overline{X}_k\right)\mathbf{L}_k^a \right\} \\ &\quad - L_{k,i}^a. \end{aligned} \tag{5}$$

In Eq. (5), $\mathscr{X}_i^{(b)}$ denotes the constellation subset of $\mathscr{X}$ with the i-th labeled bit $b_i = b \in \{0,1\}$, and the m-length vector $\mathbf{s}\left(\overline{X}_k\right) = \mathbf{1} - 2\mathbf{b}\left(\overline{X}_k\right)$, where $\overline{X}_k$ is the candidate symbol on the k-th subcarrier and $\mathbf{b}\left(\overline{X}_k\right)$ represents the associated labeled bit vector of $\overline{X}_k$. The de-interleaved demapping LLR $\mathbf{L}^{a,d}$ is then fed to soft decoder to calculate the *a posteriori* LLR $\mathbf{L}^{p,d}$ for data recovery and extrinsic LLR $\mathbf{L}^{e,d}$ for iterative demapping.

3. Enhanced iterative demapping algorithm

In this section, a novel extrinsic LLR criterion derivation for clipped DCO-OFDM is proposed to generate more accurate LLR by incorporating the correlation between clipping distortions over different subcarriers. Besides that, a simplified demapping algorithm is incorporated in the proposed demapper to reduce the complexity of extrinsic LLR calculation.

3.1. *Novel LLR calculation criterion*

In the conventional demapper of BICM-ID scheme, LLR vector on each subcarrier is separately calculated according to Max-Log-MAP algorithm and the correlation between clipping distortions over different subcarriers is neglected, which impairs the performance of BICM-ID based VLC systems. Hence, symbol vector on occupied subcarriers should be exploited to recover the clipping distortion and generate more accurate LLR. Assuming that $\overline{\mathbf{X}}$ is arbitrary N_{used}-length candidate symbol vector, the likelihood of received signal $\mathbf{y}$ is formulated as

$$P\left(\mathbf{y}\,|\overline{\mathbf{X}}\right) = \frac{1}{\left(2\pi\sigma^2\right)^{\frac{N}{2}}} \exp\left(-\frac{1}{2\sigma^2}\left\|\left\lfloor \mathbf{x}^r\left(\overline{\mathbf{X}}\right)\right\rfloor - \mathbf{y}\right\|^2\right). \tag{6}$$

The reconstructed time-domain DCO-OFDM signal $\mathbf{x}^r\left(\overline{\mathbf{X}}\right)$ is acquired as

$$\mathbf{x}^r\left(\overline{\mathbf{X}}\right) = IFFT\left(HS\left(\overline{\mathbf{X}}\right)\right) + B_{\mathrm{DC}}. \tag{7}$$

Since the clipping operation damages the orthogonality of subcarriers, the calculation of $L_{k,i}^a$ requires to search all the candidate symbol vector $\overline{\mathbf{X}}$ rather than the candidate symbol $\overline{X}_k$, which is only the k-th element of $\overline{\mathbf{X}}$, to maximize the *a posteriori* probability. Given that $\mathscr{X}_{k,i}^{N_{used},(b)}$

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denotes the subset of $\mathscr{X}^{N_{used}}$, where the i-th labeled bit of the k symbol is fixed as $b \in \{0,1\}$, by substituting single symbol with symbol vector in Eq. (5), we have

$$
\begin{aligned}
L_{k,i}^{e} &\approx \log \frac{\max\limits_{\overline{X} \in \mathscr{X}_{k,i}^{N_{used},(0)}} \left\{ P\left(\mathbf{y}|\overline{\mathbf{X}}\right) P\left(\overline{\mathbf{X}}|\mathbf{L}^{a}\right) \right\}}{\max\limits_{\overline{X} \in \mathscr{X}_{k,i}^{N_{used},(1)}} \left\{ P\left(\mathbf{y}|\overline{\mathbf{X}}\right) P\left(\overline{\mathbf{X}}|\mathbf{L}^{a}\right) \right\}} - L_{k,i}^{a} \\
&= \max_{\overline{\mathbf{X}} \in \mathscr{X}_{k,i}^{N_{used},(0)}} \left\{ -\frac{1}{2\sigma^{2}} \left\| \left\lfloor \mathbf{x}^{r}\left(\overline{\mathbf{X}}\right) \right\rfloor - \mathbf{y} \right\|^{2} + \frac{1}{2} \mathbf{s}^{\mathrm{T}}\left(\overline{\mathbf{X}}\right) \mathbf{L}^{a} \right\} \\
&- \max_{\overline{\mathbf{X}} \in \mathscr{X}_{k,i}^{N_{used},(1)}} \left\{ -\frac{1}{2\sigma^{2}} \left\| \left\lfloor \mathbf{x}^{r}\left(\overline{\mathbf{X}}\right) \right\rfloor - \mathbf{y} \right\|^{2} + \frac{1}{2} \mathbf{s}^{\mathrm{T}}\left(\overline{\mathbf{X}}\right) \mathbf{L}^{a} \right\} - L_{k,i}^{a}.
\end{aligned} \tag{8}
$$

It is obvious that the proposed LLR calculation criterion in Eq. (8) consumes N multiplication to acquire Euclidean distance $\left\| \left\lfloor \mathbf{x}^{r}\left(\overline{\mathbf{X}}\right) \right\rfloor - \mathbf{y} \right\|^{2}$ and searches all $2^{mN_{used}}$ possible candidate vector $\overline{\mathbf{X}}$ to find the maximum *a posteriori* probability. Hence, Eq. (8) exhausts $2^{mN_{used}}$ IFFT operation and $2^{mN_{used}}N$ multiplication to calculate each $L_{k,i}^{e}$, which makes the proposed demapping algorithm infeasible in practice.

3.2. Simplified demapping algorithm

In order to reduce the complexity of the proposed LLR criterion, three steps are presented to simplify the maximization operation, Euclidean distance calculation and time-domain signal regeneration.

3.2.1. Target vector subset narrowing

It is reasonable to assume that the transmitted symbol vector $\mathbf{X}$ approaches the optimal solution of maximizing *a posteriori* probability in log-domain when high SNR channel is adopted. An accurate estimated symbol vector $\mathbf{X}^{a}\left(\mathbf{L}^{a}\right) = \left[X^{a}\left(\mathbf{L}_{1}^{a}\right), X^{a}\left(\mathbf{L}_{2}^{a}\right), \ldots, X^{a}\left(\mathbf{L}_{N_{used}}^{a}\right)\right]^{\mathrm{T}}$ could be simply acquired based on *a priori* LLR $\mathbf{L}^{a}$ provided by the decoder. The associated LLR vector $\mathbf{L}_{k}^{a}$ on k-th subcarrier is transferred into *a priori* bit vector $\mathbf{b}^{a}\left(\mathbf{L}_{k}^{a}\right)$ as

$$
b_{i}^{k}\left(\mathbf{L}_{k}^{a}\right) = \begin{cases} 0, L_{k,i}^{a} \geqslant 0 \\ 1, L_{k,i}^{a} < 0 \end{cases}, \tag{9}
$$

where $b_{i}^{k}\left(\mathbf{L}_{k}^{a}\right)$ is the i-th element of $\mathbf{b}^{a}\left(\mathbf{L}_{k}^{a}\right)$. According to the constellation set $\mathscr{X}$, $\mathbf{b}^{a}\left(\mathbf{L}_{k}^{a}\right)$ is mapped into estimated symbol $X^{a}\left(\mathbf{L}_{k}^{a}\right)$. With the aid of $\mathbf{X}^{a}\left(\mathbf{L}^{a}\right)$, a simplified strategy is proposed to narrow the target vector subset $\mathscr{X}_{k,i}^{N_{used},(b)}, b \in \{0,1\}$ in Eq. (8). When the extrinsic LLR $L_{k,i}^{e}$ of the k-th symbol is calculated, candidate symbol vector $\overline{\mathbf{X}} \in \mathscr{X}_{k,i}^{N_{used},(b)}, b \in \{0,1\}$ is substituted with $\overline{\mathbf{X}}^{k}\left(\overline{X}_{k}, \mathbf{L}^{a}\right) = \left[\overline{X}_{1}^{k}\left(\overline{X}_{k}, \mathbf{L}^{a}\right), \overline{X}_{2}^{k}\left(\overline{X}_{k}, \mathbf{L}^{a}\right), \ldots \overline{X}_{N_{used}}^{k}\left(\overline{X}_{k}, \mathbf{L}^{a}\right)\right]^{\mathrm{T}}$ which is assigned as

$$
\overline{X}_{l}^{k}\left(\overline{X}, \mathbf{L}^{a}\right) = \begin{cases} \overline{X}_{k} & , l = k \\ X^{a}\left(\mathbf{L}_{l}^{a}\right) & , l \neq k \end{cases}, \tag{10}
$$

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where $\overline{X}_k \in \mathscr{X}_i^{(b)}, b \in \{0,1\}$. By incorporating this approximation, the LLR criterion for clipped DCO-OFDM signals in Eq. (8) is simplified as

$$\begin{aligned} L_{k,i}^e = & \max_{\overline{X}_k \in \mathscr{X}_i^{(0)}} \left\{ -\frac{1}{2\sigma^2} \left\| \left\lfloor \mathbf{x}^r \left(\overline{\mathbf{X}}^k \left(\overline{X}_k, \mathbf{L}^a \right) \right) \right\rfloor - \mathbf{y} \right\|^2 + \frac{1}{2} \mathbf{s}^{\mathrm{T}} \left(\overline{X}_k \right) \mathbf{L}_k^a \right\} \\ & - \max_{\overline{X}_k \in \mathscr{X}_i^{(1)}} \left\{ -\frac{1}{2\sigma^2} \left\| \left\lfloor \mathbf{x}^r \left(\overline{\mathbf{X}}^k \left(\overline{X}_k, \mathbf{L}^a \right) \right) \right\rfloor - \mathbf{y} \right\|^2 + \frac{1}{2} \mathbf{s}^{\mathrm{T}} \left(\overline{X}_k \right) \mathbf{L}_k^a \right\} - L_{k,i}^a. \end{aligned} \tag{11}$$

Thus, the essential number of searching possible vectors is reduced from $2^{mN_{used}}$ to 2^m.

3.2.2. One-dimension Euclidean distance

Euclidean distance is repeatedly calculated in maximization operation of Eq. (11), which consumes extremely high complexity. In order to further reduce the demapping complexity, N-dimension time-domain Euclidean distance could be approximated as one-dimension frequency-domain counterpart. According to Parseval's theorem, the powers of time-domain and frequency-domain signals are equivalent, we have

$$\left\| \left\lfloor \mathbf{x}^r \left(\overline{\mathbf{X}}^k \left(\overline{X}_k, \mathbf{L}^a \right) \right) \right\rfloor - \mathbf{y} \right\|^2 = \sum_{l=0}^{N} \left\| X_l^c \left(\overline{X}_k, \mathbf{L}^a \right) - Y_l \right\|^2, \tag{12}$$

where $X_l^c \left(\overline{X}_k, \mathbf{L}^a \right)$ represents the l-th frequency-domain symbol of clipped regenerated time-domain DCO-OFDM signal $\left\lfloor \mathbf{x}^r \left(\overline{\mathbf{X}}^k \left(\overline{X}_k, \mathbf{L}^a \right) \right) \right\rfloor$. Due to simplification in step 1), two candidate symbol vectors $\overline{\mathbf{X}}^k \left(\overline{X}_k, \mathbf{L}^a \right)$ in Eq. (11) share the same frequency-domain symbols on all subcarriers except the k-th element and the clipping distortions of the regenerated DCO-OFDM signals differ slightly since only the candidate symbol $\overline{X}_k$ on the k-th subcarrier is changed. Hence, it is reasonable to approximate that $X_l^c \left(\overline{X}_k, \mathbf{L}^a \right), l \neq k$ or $l \neq N-k$ are non-correlated with $\overline{X}_k$ and could be neglected in calculation of the difference between two maximum *a posteriori* probability in log-domain. By substituting Eq. (12) in Eq. (11), we have

$$\begin{aligned} L_{k,i}^e = & \max_{\overline{X}_k \in \mathscr{X}_i^{(0)}} \left\{ -\frac{1}{2\sigma^2} \sum_{l=0}^{N} \left\| X_l^c \left(\overline{X}_k, \mathbf{L}^a \right) - Y_l \right\|^2 + \frac{1}{2} \mathbf{s}^{\mathrm{T}} \left(\overline{X}_k \right) \mathbf{L}_k^a \right\} \\ & - \max_{\overline{X}_k \in \mathscr{X}_i^{(1)}} \left\{ -\frac{1}{2\sigma^2} \sum_{l=0}^{N} \left\| X_l^c \left(\overline{X}_k, \mathbf{L}^a \right) - Y_l \right\|^2 + \frac{1}{2} \mathbf{s}^{\mathrm{T}} \left(\overline{X}_k \right) \mathbf{L}_k^a \right\} - L_{k,i}^a \\ \approx & \max_{\overline{X}_k \in \mathscr{X}_i^{(0)}} \left\{ -\frac{1}{\sigma^2} \left\| X_k^c \left(\overline{X}_k, \mathbf{L}^a \right) - Y_k \right\|^2 + \frac{1}{2} \mathbf{s}^{\mathrm{T}} \left(\overline{X}_k \right) \mathbf{L}_k^a \right\} \\ & - \max_{\overline{X}_k \in \mathscr{X}_i^{(1)}} \left\{ -\frac{1}{\sigma^2} \left\| X_k^c \left(\overline{X}_k, \mathbf{L}^a \right) - Y_k \right\|^2 + \frac{1}{2} \mathbf{s}^{\mathrm{T}} \left(\overline{X}_k \right) \mathbf{L}_k^a \right\} - L_{k,i}^a, \end{aligned} \tag{13}$$

where the extrinsic LLR for clipped DCO-OFDM is acquired by only calculating one-dimension Euclidean distance. It is worthwhile to note that the k-th frequency-domain symbol $X_l^c \left(\overline{X}_k, \mathbf{L}^a \right)$ is simply calculated as

$$X_k^c \left(\overline{X}_k, \mathbf{L}^a \right) = \sqrt{N} \mathbf{e}_k^{\mathrm{T}} \left\lfloor \mathbf{x}^r \left(\overline{\mathbf{X}}^k \left(\overline{X}_k, \mathbf{L}^a \right) \right) \right\rfloor, \tag{14}$$

where $\mathbf{e}_k = \left[\exp\left(\frac{2\pi}{N} 0k \right), \exp\left(\frac{2\pi}{N} 1k \right), \ldots, \exp\left(\frac{2\pi}{N} (N-1) k \right) \right]^{\mathrm{T}}$. The above transformation involves phase shifting and the multiplication is not necessary.

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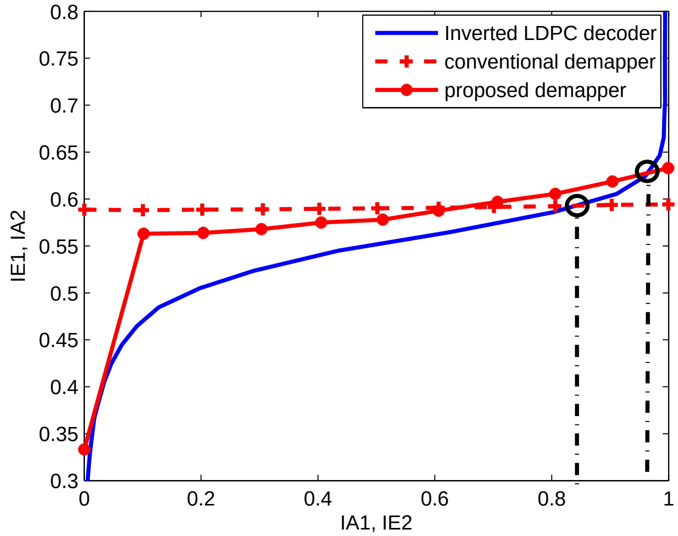

Fig. 2. The EXIT chart of the conventional and the proposed demapper for 16QAM and SNR = 8.5 dB.

3.2.3. Simplified time-domain signal regeneration

There exists only one different frequency-domain symbol on the k-th subcarrier between the estimated vector $\mathbf{X}^a(\mathbf{L}^a)$ and the candidate vector $\overline{\mathbf{X}}^k(\overline{X}_k, \mathbf{L}^a)$. This property could be exploited to reduce the number of IFFT operation significantly in time-domain signal regeneration. Initially, $\mathbf{X}^a(\mathbf{L}^a)$ is given in step 1) and transferred into associated DCO-OFDM time-domain signal $\mathbf{x}^r(\mathbf{X}^a(\mathbf{L}^a))$ according to Eq. (7), which consume only one IFFT operation. When each extrinsic LLR $L^e_{k,i}$ is processed with the simplified demapping algorithm in Eq. (13), the regenerated DCO-OFDM time-domain signal $\mathbf{x}^r\left(\overline{\mathbf{X}}^k(\overline{X}_k, \mathbf{L}^a)\right)$ is acquired as

$$x_n^r\left(\overline{\mathbf{X}}^k(\overline{X}_k, \mathbf{L}^a)\right) = x_n^r(\mathbf{X}^a(\mathbf{L}^a)) + \frac{2}{\sqrt{N}}\mathrm{Re}\left(\left(\overline{X}_k - X^a(\mathbf{L}^a_k)\right)\exp\left(\frac{2\pi}{N}kn\right)\right), \quad n = 0, 1, \ldots, N-1, \tag{15}$$

where $x_n^r\left(\overline{\mathbf{X}}^k(\overline{X}, \mathbf{L}^a)\right)$ and $x_n^r(\mathbf{X}^a(\mathbf{L}^a))$ are the n-th elements of $\mathbf{x}^r\left(\overline{\mathbf{X}}^k(\overline{X}, \mathbf{L}^a)\right)$ and $\mathbf{x}^r(\mathbf{X}^a(\mathbf{L}^a))$ respectively. Hence, no extra IFFT operation is required in the simplified demapping algorithm.

In summary, the proposed demapping algorithm consumes 2^m multiplication and no extra IFFT operation when calculating each extrinsic LLR $L^e_{k,i}$ and could reduce the implementation complexity significantly. Compared with the conventional demapper based on Eq. (5), the proposed demapper consumes one additional FFT module to process each DCO-OFDM block and requires the same number of multiplications to calculate each extrinsic LLR.

4. Simulation results

In this section, both extrinsic information transfer (EXIT) chart [16] analysis and bit error rate (BER) curves are presented to investigate the performance of the proposed iterative demapper

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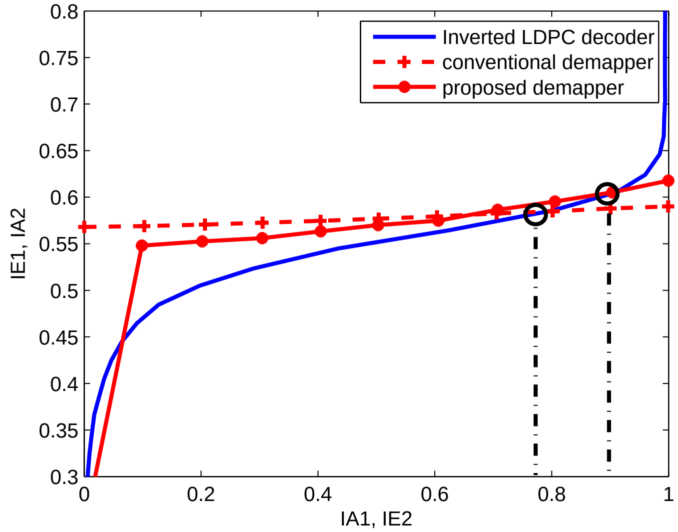

Fig. 3. The EXIT chart of the conventional and the proposed demapper for 64QAM and SNR = 12 dB.

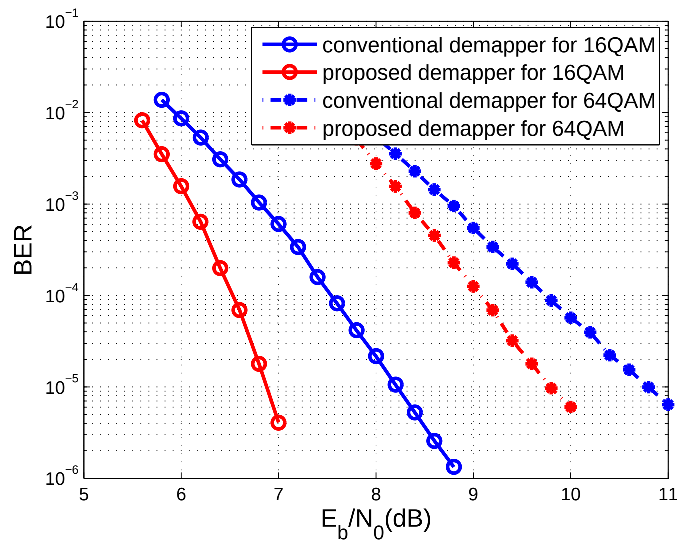

Fig. 4. The BER performance of the conventional and the proposed demapper for clipped DCO-OFDM VLC systems.

for clipped DCO-OFDM. The numbers of total and occupied subcarrier were set as $N = 64$ and $N_{used} = 27$ respectively. 16-ary and 64-ary quadrature amplitude modulation (QAM) constellations were adopted. Besides that, 1/2-rate 1296-bit low density parity check (LDPC) code from IEEE 802.11 specification [17] was selected. Since the row and column of LDPC code

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check matrix were interleaved, interleaver was omitted in simulations. In the doubled-side clipping of Eq. (4), the DC bias voltage B_{DC} was set as $(A_{\min}+A_{\max})/2$ to make full use of the dynamic region [18] and the clipping levels were set as 9 dB, 11 dB for 16QAM and 64QAM, respectively.

Figure 2 and Fig. 3 illustrate the EXIT charts of the conventional, the proposed demapper and LDPC decoder over AWGN channel with SNR = 8.5 dB and SNR = 12 dB for 16QAM and 64QAM, respectively. As for both figures, when the mutual information between $\mathbf{L}^a$ and transmitted bits equals to zero, the output mutual information of LLR $\mathbf{L}^e$ calculated by the proposed demapper is much lower than the conventional demapper. This phenomenon implies that in initial iteration where the *a priori* LLR $\mathbf{L}^a$ is absent, the proposed demapper could not generate estimated vector $\mathbf{X}^a(\mathbf{L}^a)$ and export inaccurate extrinsic LLR. Thus, the conventional Max-Log-MAP algorithm should be adopted in the first iteration of the proposed demapper. However, as the mutual information between $\mathbf{L}^a$ and the transmitted bits increases in Fig. 2, the EXIT transfer curve of the proposed demapper exhibits larger slope and converges to the point where mutual information of the decoder output LLR equals to 0.96. In contrast, the conventional demapper only reaches the point of 0.84. The similar results are presented in the case of high order modulation 64QAM. As depicted in Fig. 3, converge points of the conventional and the proposed demapper are 0.79 and 0.91, respectively. The EXIT chart analysis proves that the proposed iterative demapper generates more accurate extrinsic LLR and exhibits the enhanced convergency for clipped DCO-OFDM.

Figure 4 presents the BER curves of clipped DCO-OFDM VLC systems based on BICM-ID scheme with different order modulation constellations and iterative demappers over the AWGN channel. The proposed iterative demapping algorithm achieves a significant BER performance gain of 1.3 dB and 1.0 dB for 16QAM and 64QAM at BER = 10^{-5} compared with its conventional counterpart, which demonstrates that the extrinsic LLR provided by the proposed demapper assists the decoder to improve bit error correction. The simulation results proves that the proposed iterative demapper is capable of enhancing the performance of the clipped DCO-OFDM VLC systems and mitigating the clipping distortions.

5. Conclusion

In this paper, a BICM-ID scheme for clipped DCO-OFDM is investigated for VLC systems. Considering the correlation between clipping distortions over different subcarriers, we derive a novel extrinsic LLR criterion for clipped DCO-OFDM to provide more accurate LLR for the decoder and develop a simplified demapping algorithm to reduce the complexity of the proposed extrinsic LLR criterion. Simulation results demonstrate that BICM-ID scheme with the proposed iterative demapper enhances the convergency and BER performance of clipped DCO-OFDM VLC systems and mitigates the LED nonlinearity effectively.

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